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Single Stage Amplifiers

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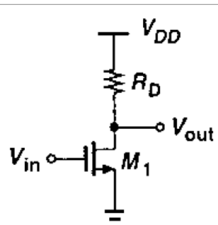
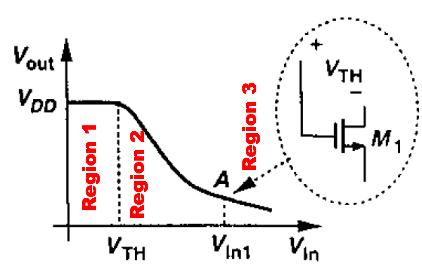
Common Source Resistor Load



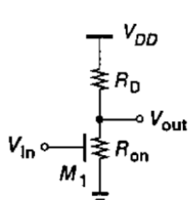


Large Signal Behavior:

Region 2: $V_{out} = V_{DD} - R_D\frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{in} - V_{TH})^2$

$V_{IN}=V_{IN1}$: $V_{in1} - V_{TH} = V_{DD} - R_D\frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{in1} - V_{TH})^2$

$$V_{out} = V_{DD} - R_D\frac{1}{2}\mu_n C_{ox}\frac{W}{L}[2(V_{in} - V_{TH})V_{out} - V_{out}^2]$$

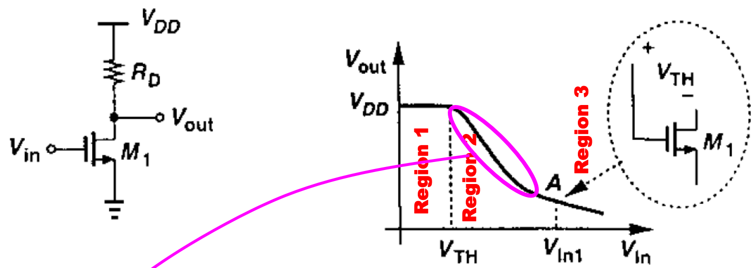


In deep triode region:

$$V_{out} = V_{DD}\frac{R_{on}}{R_{on}+R_D}$$

$$V_{out} = \frac{V_{DD}}{1+\mu_n C_{ox}\frac{W}{L}R_D(V_{in}-V_{TH})}$$

Common Source Resistor Load



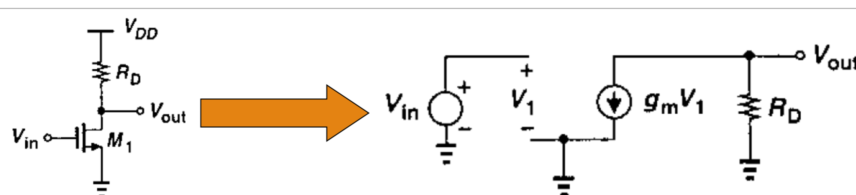
Gain is slope:

$$A_v = \frac{\partial V_{out}}{\partial V_{in}}$$

$$A_v = -R_D \mu_n C_{ox} \frac{W}{L}(V_{in} - V_{TH})$$

$$A_v = -g_m R_D$$

Common Source Resistor Load



- Small Signal View Gain= $V_{out} = -g_m V_1 R_D = -g_m V_{in} R_D$
- g_m varies with the input signal as: $g_m = \mu_n C_{ox} \frac{W}{L}(V_{GS} - V_{TH})$
- Gain varies with input signal.
- How to maximize gain of a resistive loaded CS stage?

$$A_v = -\sqrt{2\mu_n C_{ox} \frac{W}{L} I_D} \frac{V_{RD}}{I_D} \quad \Rightarrow \quad A_v = -\sqrt{2\mu_n C_{ox} \frac{W}{L}} \frac{V_{RD}}{\sqrt{I_D}}$$

- Av can be increased by increasing *W/L*, increasing V_{RD}, or decreasing I_D.
- Implications of each:
  - Increasing *W/L* causes increasing in capacitance.
  - Increasing V_{RD} causes reduced signal swing.
  - Reducing I_D should result in increase in R_D to have same Output swing ➔ higher time constant ➔ slow circuit.

Common Source Resistor Load

- Effect of channel length modulation:

$$V_{out} = V_{DD} - R_D \tfrac{1}{2}\mu_n C_{ox}\tfrac{W}{L}(V_{in} - V_{TH})^2(1+\lambda V_{out})$$

- Gain:

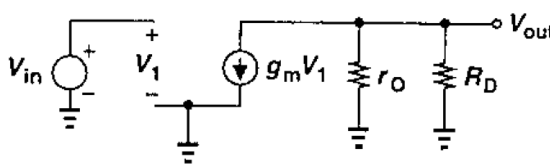
$$\frac{\partial V_{out}}{\partial V_{in}} = -R_D \mu_n C_{ox}\frac{W}{L}(V_{in} - V_{TH})(1+\lambda V_{out}) - R_D \tfrac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{in} - V_{TH})^2 \lambda \frac{\partial V_{out}}{\partial V_{in}}$$

$$\implies A_v = -R_D g_m - R_D I_D \lambda A_v$$

$$\implies A_v = -\frac{g_m R_D}{1 + R_D \lambda I_D}$$

$$\implies A_v = -g_m \frac{r_o R_D}{r_o + R_D} = -g_m (r_o || R_D)$$

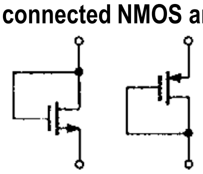
- Small signal model with channel length modulation:

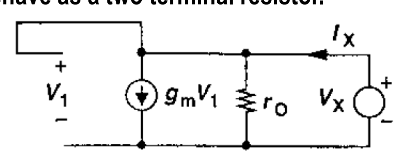


$$\implies A_v = -g_m \frac{r_o R_D}{r_o + R_D}$$

Diode Connected Transistor

- Diode connected NMOS and PMOS behave as a two terminal resistor.

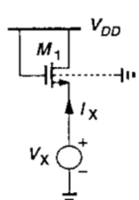


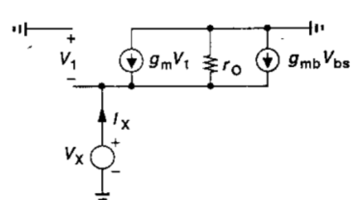


- The impedance across the two terminals is obtained by solving KCL: $I_X = \frac{V_X}{r_o} + g_m V_X$
- In presence of body-effect: $(g_m + g_{mb})V_X + \frac{V_X}{r_o} = I_X$

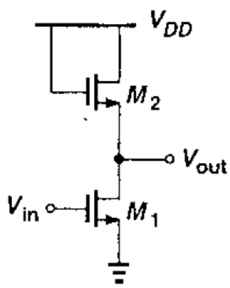
$$\implies \frac{V_X}{I_X} = \frac{1}{g_m + g_{mb} + r_o^{-1}} = \frac{1}{g_m + g_{mb}} || r_o \approx \frac{1}{g_m + g_{mb}} = \frac{1}{g_m} || r_o || \frac{1}{g_{mb}}$$

Why is this lower than without body effect??





CS Diode Connected Load



$$A_v = -g_{m1}\frac{1}{g_{m2}+g_{mb2}} = -\frac{g_{m1}}{g_{m2}}\frac{1}{1+\eta} \quad \text{where } \eta = \frac{g_{mb2}}{g_{m2}}$$

$$A_v = -\frac{\sqrt{2\mu_n C_{ox}(W/L)_1 I_{D1}}}{\sqrt{2\mu_n C_{ox}(W/L)_2 I_{D2}}}\frac{1}{1+\eta}$$

- Since $I_{D1}=I_{D2}$:

$$\Longrightarrow A_v = -\sqrt{\frac{(W/L)_1}{(W/L)_2}}\frac{1}{1+\eta}$$

- Gain is independent of bias current and voltage so long as M_1 is in saturation. (even in large signal sense)

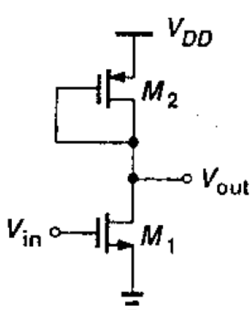
LARGE SIGNAL BEHAVIOR: $\frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)_1 (V_{in}-V_{TH1})^2 = \frac{1}{2}\mu_n C_{ox}\left(\frac{W}{L}\right)_2 (V_{DD}-V_{out}-V_{TH2})^2$

$$\Longrightarrow \sqrt{\left(\frac{W}{L}\right)_1}(V_{in}-V_{TH1}) = \sqrt{\left(\frac{W}{L}\right)_2}(V_{DD}-V_{out}-V_{TH2})$$

Differentiating both sides wrt V_{in} we get

$$\sqrt{\left(\frac{W}{L}\right)_1} = \sqrt{\left(\frac{W}{L}\right)_2}\left(-\frac{\partial V_{out}}{\partial V_{in}} - \frac{\partial V_{TH2}}{\partial V_{in}}\right) \Longrightarrow \frac{\partial V_{out}}{\partial Vin} = -\sqrt{\frac{(W/L)_1}{(W/L)_2}}\frac{1}{1+\eta}$$

CS Diode Connected Load



- No body effect: $A_v = -\sqrt{\frac{\mu_n (W/L)_1}{\mu_p (W/L)_2}}$
- Gain is a function of ratio of the widths of transistors.
- For higher gain ➔ big input devices and small load devices.
- For a particular current we have $I_{D1}=I_{D2}$ we have,

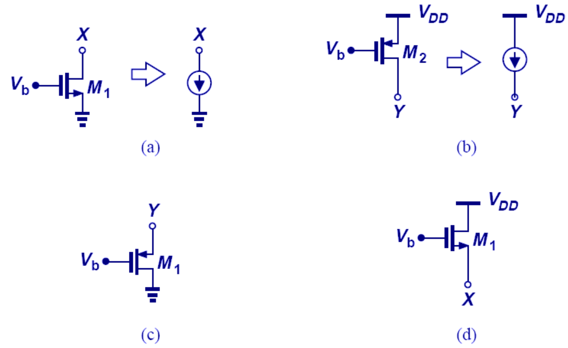
$$\mu_n\left(\frac{W}{L}\right)_1 (V_{GS1}-V_{TH1})^2 = \mu_p\left(\frac{W}{L}\right)_2 (V_{GS2}-V_{TH2})^2$$

$$\frac{|V_{GS2}-V_{TH2}|}{V_{GS1}-V_{TH1}} = A_v$$

- Output is limited to maximum of (V_{DD}-$|V_{TH}|$).
- In nanometer CMOS channel-length modulation is significant:

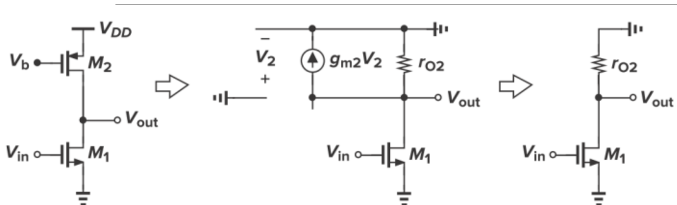
$$A_v = -g_{m1}\left(\frac{1}{g_{m2}}||r_{o1}||r_{o2}\right)$$

Current Sources



- When in saturation region, a MOSFET behaves as a current source.
- NMOS draws current from a point to ground (sinks current), whereas PMOS draws current from V_{DD} to a point (sources current).

CS Current Source Load



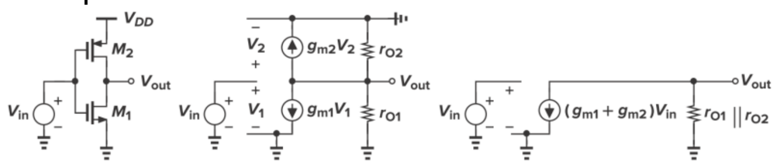
$$A_v = -g_{m1}(r_{o1}||r_{o2})$$

- Larger swing at output.
- Output voltage is not well defined ➔ need some kind of feedback to define it (more later).

- Gain is proportional to output impedance. $g_{m1} r_{o1} = \sqrt{2\left(\frac{W}{L}\right)_1 \mu_n C_{ox} I_D \frac{1}{\lambda I_D}}$
- $\lambda \propto 1/L$
- $r_O \approx 1/\lambda I_D$ ➔ $r_O \propto L/I_D$
- To increase gain if L_1 is increased then W_1 needs to be increased otherwise higher overdrive is needed ➔ reducing output swing➔ not desirable.
- Also $g_{m1} \propto \sqrt{(W/L)_1}$ not scaling W_1 will reduce g_m ➔ gain.

CS Current Source Load

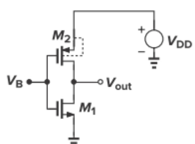
- **Inverter Based CS amplifer:**



$$-(g_{m1}+g_{m2})V_{in}(r_{o1}||r_{o2}) = V_{out}$$

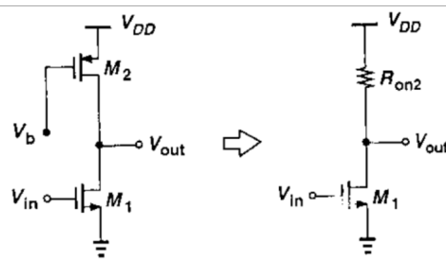
$$A_v = -(g_{m1}+g_{m2})(r_{o1}||r_{o2})$$

- **Suffers from poorly defined bias point.**
- **Suffers from supply noise amplification (poor PSRR).**



$$\frac{V_{out}}{V_{DD}} = \frac{g_{m2}r_{o2}+1}{r_{o2}+r_{o1}}r_{o1} = \left(g_{m2}+\frac{1}{r_{o2}}\right)(r_{o1}||r_{o2})$$

CS Triode Load



- Gain = $g_m R_{on2}$ and $R_{on2} = \frac{1}{\mu_p C_{ox}(W/L)_2(V_{DD}-V_b-|V_{THP}|)}$
- Triode load consume less headroom than diode connected load ➔ $V_{out,max} \approx V_{DD}$

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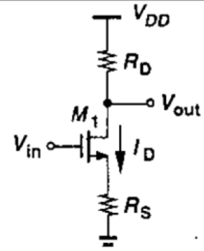
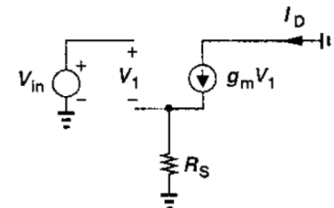
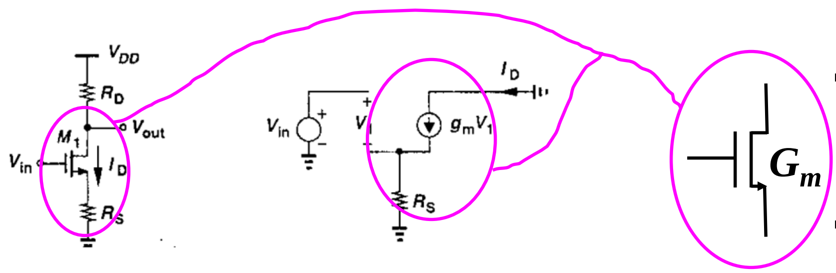
CS With Source Degeneration

- $V_{OUT} = I_D R_D$
- I_D varies nonlinearly with input voltage.
- Can we make I_D a weak function of input voltage i.e. make g_m a weak function of input voltage?







$$G_m = \frac{\partial I_D}{\partial V_{in}} = \frac{\partial f}{\partial V_{GS}} \frac{\partial V_{GS}}{\partial V_{in}}$$

- Since, $V_{GS} = V_{in} - I_D R_S$ we have $\frac{\partial V_{GS}}{\partial V_{in}} = 1 - R_S \frac{\partial I_D}{\partial V_{in}}$, thus

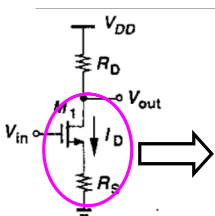
$$G_m = \left(1 - R_S \frac{\partial I_D}{\partial V_{in}}\right) \frac{\partial f}{\partial V_{GS}}$$

- But, $\frac{\partial f}{\partial V_{GS}}$ is g_m of M_1. Thus,

$$G_m = \frac{g_m}{1 + g_m R_S}$$

$$\Rightarrow A_v = -G_m R_D = -\frac{g_m R_D}{1 + g_m R_S}$$

CS With Source Degeneration

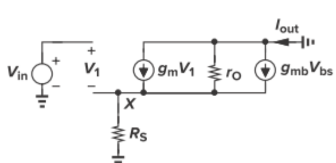


$$G_m = \frac{g_m}{1 + g_m R_S}$$

If $g_m R_S >> 1 \Rightarrow G_m = 1/R_S$

independent of input voltage.

- Small signal analysis will also give the same result. Let's analyze with body effect:



$$I_{out} = g_m V_1 - g_{mb} V_X - \frac{I_{out} R_S}{r_o}$$

$$= g_m (V_{in} - I_{out} R_S) + g_{mb}(-I_{out} R_S) - \frac{I_{out} R_S}{r_o}$$

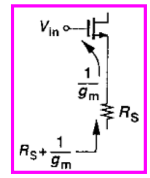
Thus, $G_m = \frac{I_{out}}{V_{in}} = \frac{g_m r_o}{R_S + [1 + (g_m + g_{mb}) R_S] r_o}$

Note: $V_{in} = V_1 + I_{out} R_S$

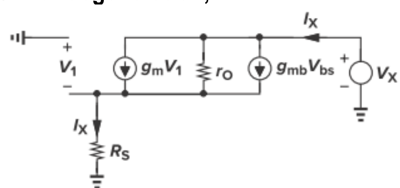
CS With Source Degeneration

- Another way to look at it , $A_v = -G_m R_D = -\frac{g_m R_D}{1+g_m R_S}$

$$\implies A_v = -\frac{R_D}{\frac{1}{g_m}+R_S}$$



- Output impedance of a CS with degeneration,



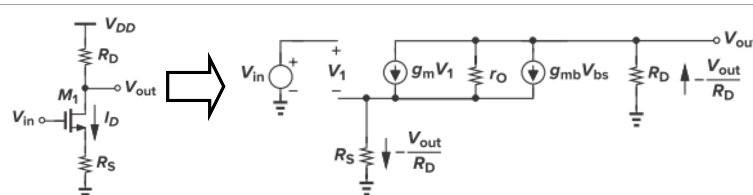
- Note that $V_1 = -I_X R_S$ and $I_X - (g_m + g_{mb})V_1 = I_X + (g_m + g_{mb})R_S I_X$
- Adding voltage drop across r_o and R_S we get,

$$r_o[I_X + (g_m + g_{mb})R_S I_X] + I_X R_S = V_X$$

$$\implies R_{out} = [1 + (g_m + g_{mb})R_S]r_o + R_S = [1 + (g_m + g_{mb})r_o]R_S + r_o$$

- This can be approximated as: $R_{out} \approx (g_m + g_{mb})r_o R_S + r_o = [1 + (g_m + g_{mb})R_S]r_o$

CS With Source Degeneration



- Solving using small-signal model incorporating everything we get:
- Step-1: $I_{ro} = -\frac{V_{out}}{R_D} - (g_m V_1 + g_{mb} V_{bs}) = -\frac{V_{out}}{R_D} - \left[g_m\left(V_{in} + V_{out}\frac{R_S}{R_D}\right) + g_{mb} V_{out}\frac{R_S}{R_D}\right]$
- Step-2: $V_{out} = I_{ro} r_o - \frac{V_{out}}{R_D} R_S$

$$= -\frac{V_{out}}{R_D} r_o - \left[g_m\left(V_{in} + V_{out}\frac{R_S}{R_D}\right) + g_{mb} V_{out}\frac{R_S}{R_D}\right] r_o - V_{out}\frac{R_S}{R_D}$$

- Step-3: $\frac{V_{out}}{V_{in}} = \frac{-g_m r_o R_D}{R_D+R_S+r_o+(g_m+g_{mb})R_S r_o}$

$$A_v = \frac{-g_m r_o R_D[R_S+r_o+(g_m+g_{mb})R_S r_o]}{R_D+R_S+r_o+(g_m+g_{mb})R_S r_o} \frac{1}{R_S+r_o+(g_m+g_{mb})R_S r_o} \qquad R_{out} \approx (g_m + g_{mb})r_o R_S$$

$$= -\frac{g_m r_o}{R_S+r_o+(g_m+g_{mb})R_S r_o} \underbrace{\frac{R_D[R_S+r_o+(g_m+g_{mb})R_S r_o]}{R_D+R_S+r_o+(g_m+g_{mb})R_S r_o}}$$

$$\implies A_v = -G_m(R_{out}||R_D)$$

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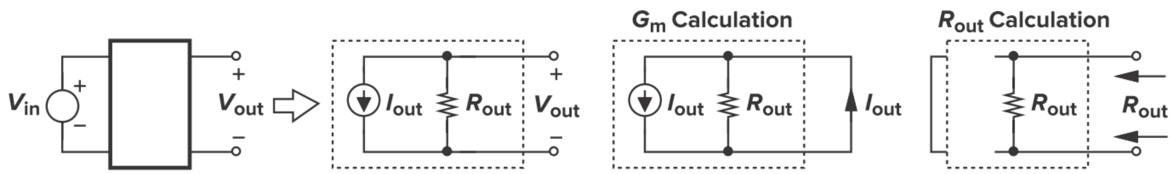
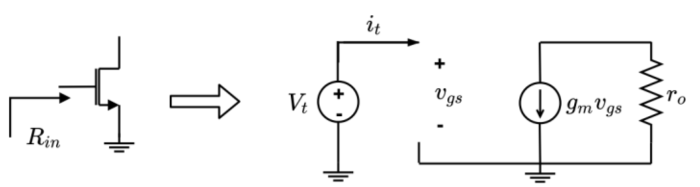
Gain Of Any Linear Circuit

- Voltage gain of a linear circuit is defined as $G_m R_{OUT}$.
  - where G_m is the equivalent transconductance of the circuit with output shorted to ground
  - R_{OUT} is the output impedance of the circuit with all independent voltage sources shorted and independent current sources opened.

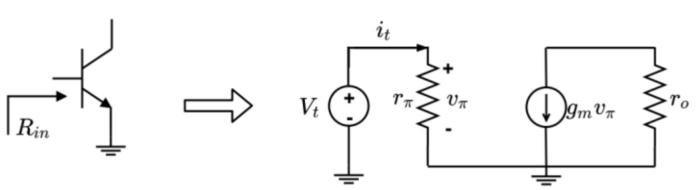


(a) (b) (c)

Resistance Looking Into Base/Gate of Transistor



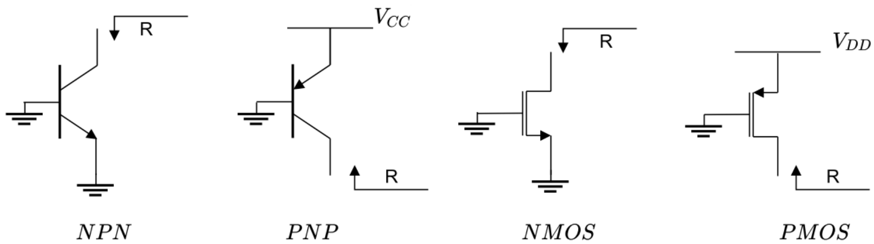
$$R_{in} = \infty$$

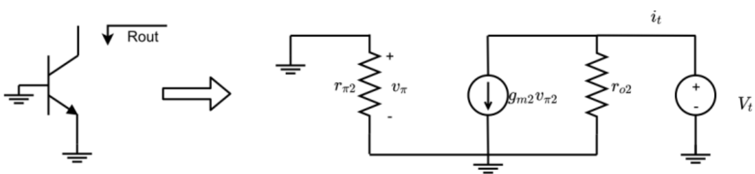


$$R_{in} = \frac{V_t}{i_t} = r_\pi$$

Note: Anything connected to drain/collector will not alter R_{in}.

Resistance Looking Into Collector/Drain of Transistor

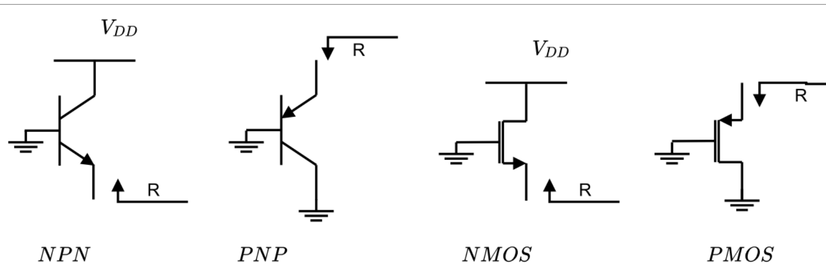


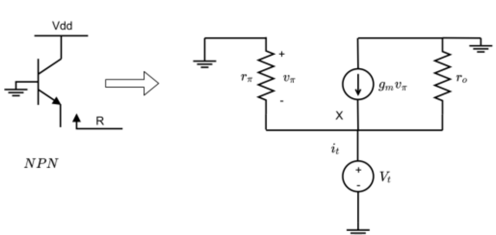


Applying KVL:

$$i_t = \frac{V_t}{r_o} \implies R_{out} = r_o$$

Resistance Looking Into Emitter/Source of Transistor





Applying KVL:

$$i_t = -g_m v_\pi + V_t/r_\pi + V_t/r_o$$

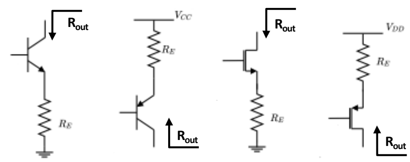
Note: $V_t = -v_\pi$

Thus,

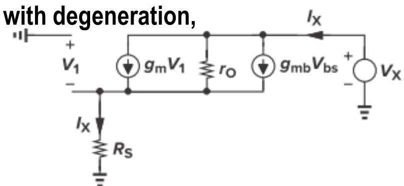
$$R = V_t/i_t = (1/g_m)||r_\pi||r_o$$

$$R \approx V_t/i_t = 1/g_m \text{ as } r_o >> (1/g_m) \text{ and } r_\pi >> (1/g_m)$$

Resistance Looking Into Emitter/Source of Degenerated Transistor



- Output impedance of a CS with degeneration,



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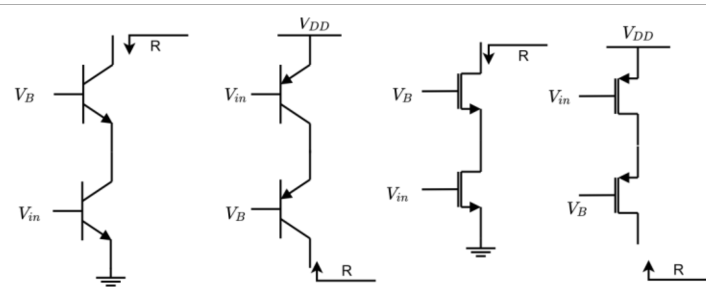
$$r_o[I_X + (g_m + g_{mb})R_S I_X] + I_X R_S = V_X$$

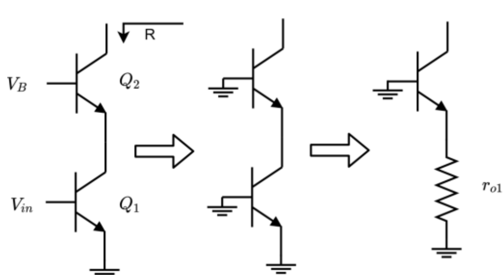
$$\implies R_{out} = [1 + (g_m + g_{mb})R_S]r_o + R_S = [1 + (g_m + g_{mb})r_o]R_S + r_o$$

- This can be approximated as: $R_{out} \approx (g_m + g_{mb})r_o R_S + r_o = [1 + (g_m + g_{mb})R_S]r_o$
- Ignoring body effect this can be approximated as:

$$R_{out} \approx g_m r_o R_S + r_o \approx g_m r_o R_S$$

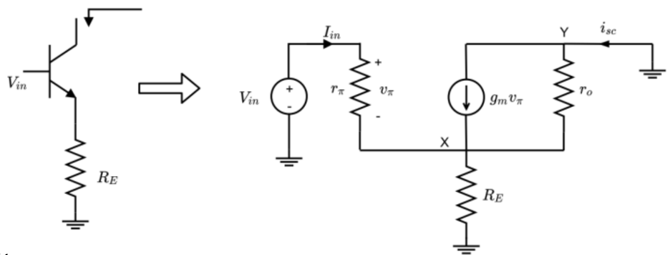
Resistance Looking Into Emitter/Source of Cascode Transistor





$$R_{out} \approx g_{m2} r_{o2} r_{o1} + r_{o2} \approx g_{m2} r_{o1} r_{o2}$$

Effective Transconductance of Degenerated Transistor



Applying KCL @ node X: $i_{sc} = -v_x/r_o + g_m(v_{in} - v_x)$

$$\implies i_{sc} = g_m v_{in} - (g_m + 1/r_o)v_x$$

Voltage across R_E at node Y: $v_x = [i_{sc} + (v_{in} - v_x)/r_\pi]R_E$

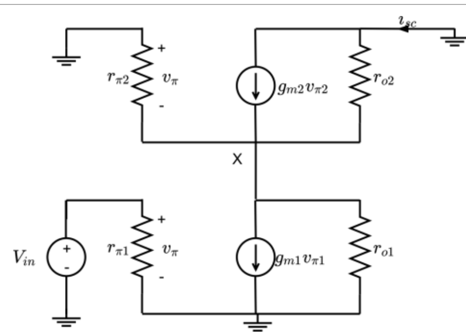
$$\implies v_x = \frac{(i_{sc}+v_{in}/r_\pi)R_E}{(1+R_E/r_\pi)}$$

Thus, substituting v_x in i_{sc} we get,

$$G_m = \frac{i_{sc}}{v_{in}} = \frac{g_m - \frac{R_E}{r_\pi r_o}}{(1+\frac{R_E}{r_\pi})+(g_m+\frac{1}{r_o})R_E}$$

In case of MOSFET this will boil down to, $G_m = \frac{i_{sc}}{v_{in}} = \frac{g_m}{1+g_m R_S}$, where R_S is the degeneration resistor.

Effective Transconductance of Cascode Transistor



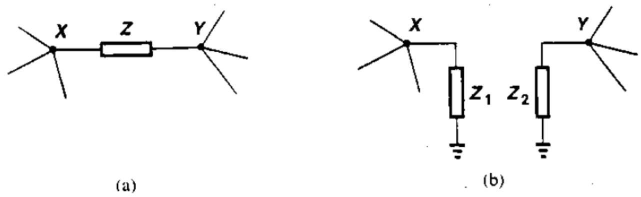
By solving the above model we get,

$$G_m = \frac{i_{sc}}{v_{in}} = \frac{g_{m1}}{1+[(\frac{1}{r_{\pi 2}}+\frac{1}{r_{o1}})(\frac{r_{o2}}{1-g_{m2}r_{o2}})]}$$

If r_{o1} and r_{o2} are large and r_π is ∞ in case of MOSFET then, $G_m = \frac{i_{sc}}{v_{in}} = g_{m1}$

Miller Effect

- Miller Theorem: If circuit in Fig.(a) is equivalent to circuit in Fig.(b) then $Z_1 = \frac{Z}{(1-A_v)}$ and $Z_2 = \frac{Z}{(1-A_v^{-1})}$, where $A_v = \frac{V_Y}{V_X}$.



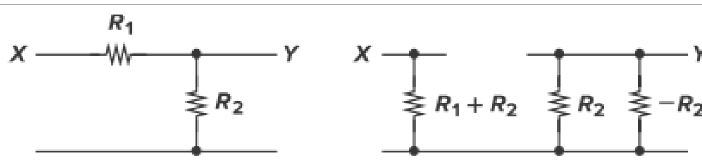
Proof: Current flowing between X and Y is equal to current flowing from X to ground in (b):

$$\frac{V_X - V_Y}{Z} = \frac{V_X}{Z_1} \Longrightarrow Z_1 = \frac{Z}{1-\frac{V_Y}{V_X}}$$

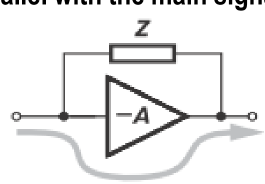
Similarly, $Z_2 = \frac{Z}{1-\frac{V_X}{V_Y}}$

If, $Z = \frac{1}{(C_F s)}$ then, $Z_1 = [1/(C_F s)]/(1 + A_v)$

Invalid application of Miller Effect

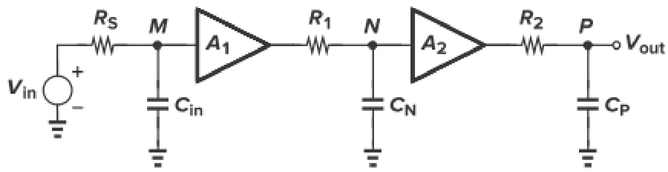


- If Z is the only path between X and Y then the conversion is invalid.
- Is valid when impedance Z is in parallel with the main signal path.



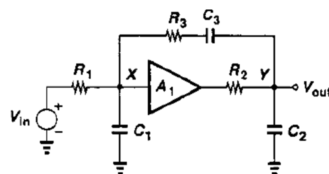
Main Signal Path

Association of Poles with Nodes

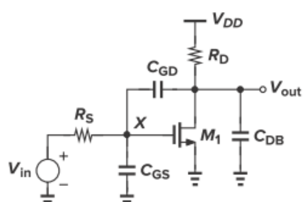


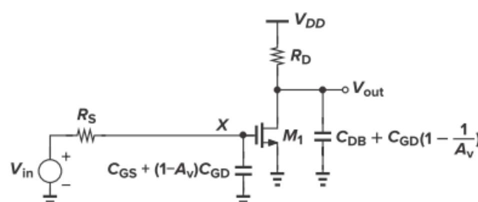
$$\frac{V_{out}}{V_{in}}(s) = \frac{A_1}{1+R_S C_{in} s} \cdot \frac{A_2}{1+R_1 C_N s} \cdot \frac{1}{1+R_2 C_P s}$$

- ❑ Interaction between nodes makes it difficult sometimes to compute the transfer function in the above straightforward manner.
- ❑ Miller effect sometimes ignores the effect of zero.



Common Source Stage

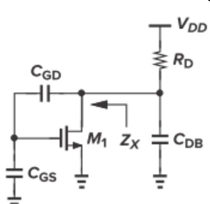




$$C_{DB} + (1 - A_v^{-1})C_{GD} \approx C_{DB} + C_{GD} \text{ (if } A_v >> 1)$$

$$\omega_{in} = \frac{1}{R_S[C_{GS}+(1+g_m R_D)C_{GD}]} \qquad \omega_{out} = \frac{1}{R_D(C_{DB}+C_{GD})}$$

- ❑ Another approx. if R_S is relatively large:



$$Z_X = \frac{1}{C_{eq}s} || \left(\frac{C_{GD}+C_{GS}}{C_{GD}} \cdot \frac{1}{g_{m1}}\right), \text{ where } C_{eq} = \frac{C_{GD}C_{GS}}{(C_{GD}+C_{GS})}$$

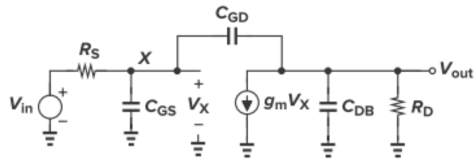
$$\omega_{out} = \frac{1}{[R_D||(\frac{C_{GD}+C_{GS}}{C_{GD}} \frac{1}{g_{m1}})](C_{eq}+C_{DB})}$$

- ❑ Thus,

$$\frac{V_{out}}{V_{in}}(s) = \frac{-g_m R_D}{(1+\frac{s}{\omega_{in}})(1+\frac{s}{\omega_{out}})}$$

Common Source Stage

- Two issues with approximating the gain as $\frac{V_{out}}{V_{in}}(s) = \frac{-g_m R_D}{(1+\frac{s}{\omega_{in}})(1+\frac{s}{\omega_{out}})}$
  - Effect of zero is ignored.
  - Gain is not $-g_m R_D$ as it is a function of the output capacitance and hence the frequency.



It is a second order system although there are 3 capacitors as three capacitors form a loop and the state of 3rd can be defined by finding the state of any two.

- Exactly solving it we get:

$$\frac{V_X - V_{in}}{R_S} + V_X C_{GS} s + (V_X - V_{out}) C_{GD} s = 0$$
$$(V_{out} - V_X) C_{GD} s + g_m V_X + V_{out}\left(\frac{1}{R_D} + C_{DB} s\right) = 0$$
$$\implies V_X = -\frac{V_{out}\left(C_{GD}s + \frac{1}{R_D} + C_{DB}s\right)}{g_m - C_{GD}s}$$
$$\implies -V_{out}\frac{[R_S^{-1} + (C_{GS}+C_{GD})s][R_D^{-1} + (C_{GD}+C_{DB})s]}{g_m - C_{GD}s} - V_{out} C_{GD} s = \frac{V_{in}}{R_S}$$
$$\implies \frac{V_{out}}{V_{in}}(s) = \frac{(C_{GD}s - g_m)R_D}{R_S R_D \xi s^2 + [R_S(1+g_m R_D)C_{GD} + R_S C_{GS} + R_D(C_{GD}+C_{DB})]s + 1}$$

where, $\xi = C_{GS}C_{GD} + C_{GS}C_{DB} + C_{GD}C_{DB}$

Common Source Stage

- Considering $\omega_{P1} << \omega_{P2}$ we can approximate the denominator as follows:

$$D = \left(\frac{s}{\omega_{p1}} + 1\right)\left(\frac{s}{\omega_{p2}} + 1\right)$$
$$= \frac{s^2}{\omega_{p1}\omega_{p2}} + \left(\frac{1}{\omega_{p1}} + \frac{1}{\omega_{p2}}\right)s + 1$$

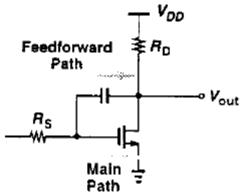
- Comparing $\omega_{in} = \frac{1}{R_S[C_{GS} + (1+g_m R_D)C_{GD}]}$ with the 1st pole of the following:

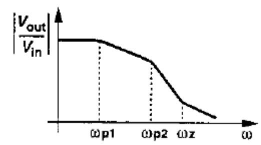
$$\omega_{p1} = \frac{1}{R_S(1+g_m R_D)C_{GD} + R_S C_{GS} + R_D(C_{GD}+C_{DB})}$$

we see that intuitive way gives very close result.

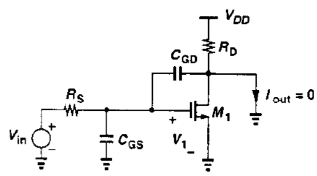
Common Source Stage

- **Presence of a zero due to feed forward path:**



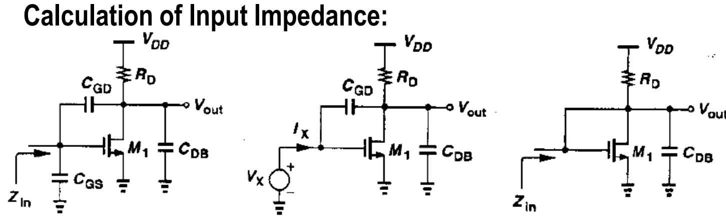


- **Computing zero:** $V_1 C_{GD} s_z = g_m V_1$ ➔ $s_z = g_m / C_{GD}$



Common Source Stage

- **Calculation of Input Impedance:**



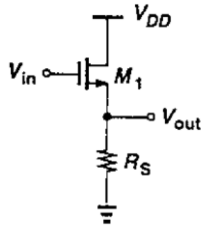
(a) (b) (c)

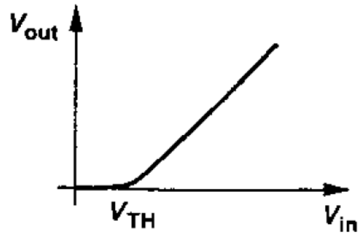
- **As 1st Order Estimate:** $Z_{in} = \frac{1}{[C_{GS}+(1+g_m R_D)C_{GD}]s}$

- **At high frequencies we get:**

$$\frac{V_X}{I_X} = \frac{1+R_D(C_{GD}+C_{DB})s}{C_{GD}s(1+g_m R_D+R_D C_{DB}s)}$$

Source Follower

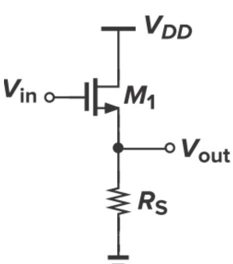




- It is also called Common-Drain.
- When $V_{in}<V_{th}$ $V_{out}=0$.
- When $V_{in} > V_{th}$, M_1 turns on and is in saturation.
- M_1 always stays in saturation.
- V_{out} follows V_{in} with a V_{GS} drop.

Source Follower

- Large signal analysis: I_D is given by $\frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{in}-V_{TH}-V_{out})^2 R_S = V_{out}$



- Differentiating V_{out} wrt V_{in} gives us the gain:

$$\frac{1}{2}\mu_n C_{ox}\frac{W}{L}2(V_{in}-V_{TH}-V_{out})\left(1-\frac{\partial V_{TH}}{\partial V_{in}}-\frac{\partial V_{out}}{\partial V_{in}}\right)R_S = \frac{\partial V_{out}}{\partial V_{in}}$$

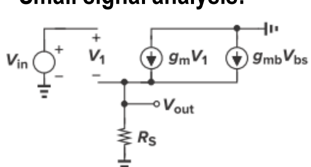
- Since, $\partial V_{TH}/\partial V_{in} = (\partial V_{TH}/\partial V_{SB})(\partial V_{SB}/\partial V_{in}) = \eta \partial V_{out}/\partial V_{in}$

$$\frac{\partial V_{out}}{\partial V_{in}} = \frac{\mu_n C_{ox}\frac{W}{L}(V_{in}-V_{TH}-V_{out})R_S}{1+\mu_n C_{ox}\frac{W}{L}(V_{in}-V_{TH}-V_{out})R_S(1+\eta)}$$

- Also, $g_m = \mu_n C_{ox}\frac{W}{L}(V_{in}-V_{TH}-V_{out})$
- Thus,

$$A_v = \frac{g_m R_S}{1+(g_m+g_{mb})R_S}$$

- Small signal analysis:



- R_{out} calculation: $I_X - g_m V_X - g_{mb} V_X = 0$

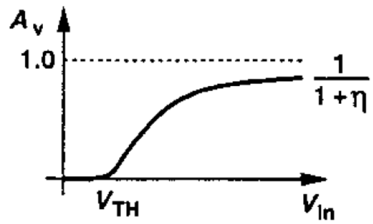
$$R_{out} = \frac{1}{g_m+g_{mb}} \Longrightarrow R_{out} = \frac{1}{g_m}||\frac{1}{g_{mb}}$$

- Gm calculation: $G_m = g_m$
- Voltage Gain: $A_v = \frac{\frac{1}{g_m}}{\frac{1}{g_m}+\frac{1}{g_{mb}}} = \frac{g_m}{g_m+g_{mb}}$

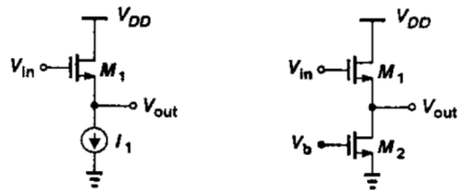
Why resistance is less with body effect?

Source Follower

$$A_v = \frac{g_m R_S}{1+(g_m+g_{mb})R_S}$$

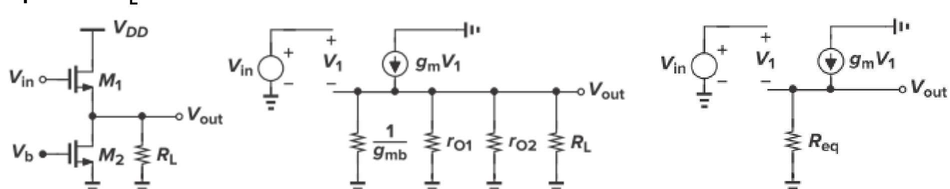


- Even if $R_S \rightarrow \infty$ $A_v < 1$.
- The drain current heavily depends on the DC voltage V_{in}.
- If I_D increases by a factor of 2, V_{GS}-V_{TH} increases by a factor of √2 ➔ huge non-linearity.
- Improve linearity by using current source instead of resistor.



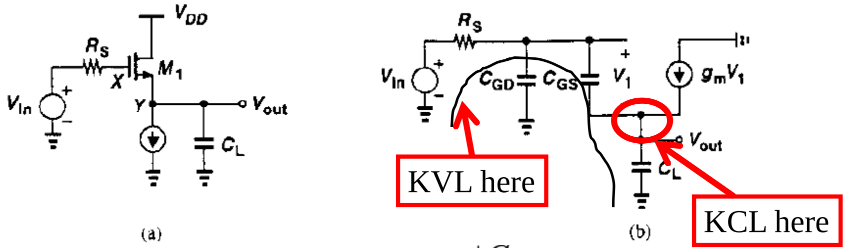
Source Follower

- Non-ideal transistor ➔ Channel length modulation.
- Driving an output load R_L.



- Noting that $R_{eq} = (1/g_{mb})||r_{o1}||r_{o2}||R_L$ we get $A_v = \frac{R_{eq}}{R_{eq}+\frac{1}{g_m}}$
- Body effect causes nonlinear change of V_{TH} with change in V_{IN} ➔ non-linearity.
- Can we avoid body effect?
  - For an NMOS ➔ difficult (Deep N-Well process can overcome body effect)
  - For a PMOS ➔ possible.
- Common uses of source follower:
  - Driving low impedance loads like 50Ω etc.
  - Level shifting of the DC level of an AC signal.

Source Follower-Frequency Response

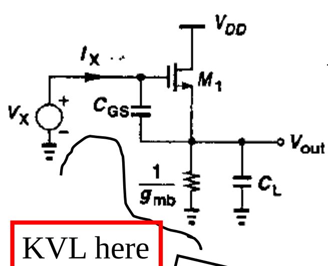


$$\frac{V_{out}}{V_{in}}(s) = \frac{g_m + C_{GS}s}{R_S(C_{GS}C_L + C_{GS}C_{GD} + C_{GD}C_L)s^2 + (g_mR_SC_{GD} + C_L + C_{GS})s + g_m}$$

LHP zero ➔ signals along C_{GS} and transistor are in phase.

- If the 2-poles are assumed to be far apart then: $\omega_{p1} \approx \frac{g_m}{g_mR_SC_{GD}+C_L+C_{GS}} = \frac{1}{R_SC_{GD}+\frac{C_L+C_{GS}}{g_m}}$
- If $R_S = 0$, then $\omega_{p1} \approx \frac{g_m}{C_L+C_{GS}}$ as expected.

Source Follower Input Impedance



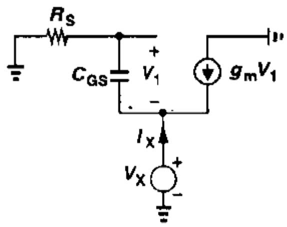
$$V_{GS} = I_X / sC_{GS} \Rightarrow I_D = g_m I_X / sC_{GS}$$

$$V_X = \frac{I_X}{C_{GS}s} + \left(I_X + \frac{g_mI_X}{C_{GS}s}\right)\left(\frac{1}{g_{mb}} || \frac{1}{C_Ls}\right)$$

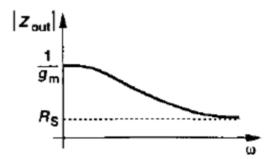
$$\Rightarrow Z_{in} = \frac{1}{C_{GS}s} + \left(1 + \frac{g_m}{C_{GS}s}\right)\frac{1}{g_{mb}+C_Ls} \Longrightarrow Z_{in} \approx \frac{1}{C_{GS}s}\left(1 + \frac{g_m}{g_{mb}}\right) + \frac{1}{g_{mb}}$$

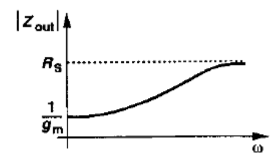
At high frequencies: $g_{mb} << C_Ls \Rightarrow Z_{in} \approx \frac{1}{C_{GS}s} + \frac{1}{C_Ls} + \frac{g_m}{C_{GS}C_Ls^2}$

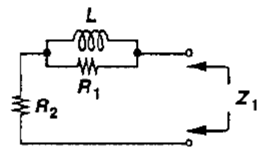
Source Follower Output Impedance



$$Z_{out} = \frac{V_X}{I_X} = \frac{R_S C_{GS} s + 1}{g_m + C_{GS} s}$$





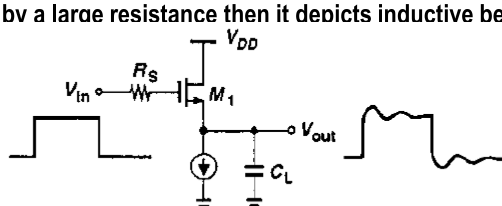


- At $\omega=0$, $Z_1=R_2=1/g_m$ and at $\omega=\infty$, $Z_1= R_S = R_1+R_2$

$$Z_{out} - \frac{1}{g_m} = \frac{C_{GS} s\left(R_S - \frac{1}{g_m}\right)}{g_m + C_{GS} s} \Rightarrow \frac{1}{Z_{out} - \frac{1}{g_m}} = \frac{1}{R_S - \frac{1}{g_m}} + \frac{1}{\frac{C_{GS} s}{g_m}\left(R_S - \frac{1}{g_m}\right)} \Rightarrow L = \frac{C_{GS}}{g_m}\left(R_S - \frac{1}{g_m}\right)$$

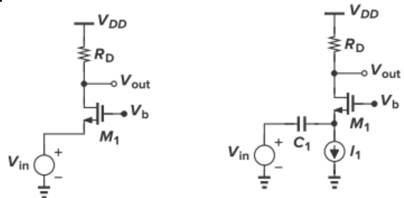
Source Follower Output Impedance As Inductive

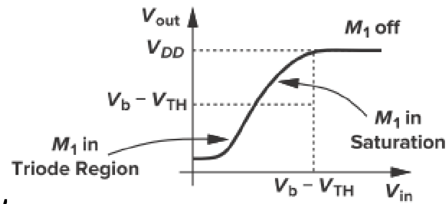
- Source follower is driven by a large resistance then it depicts inductive behavior.



Common Gate

- Signal is applied to the source:





- Large signal behavior: Lets analyze with V_{in} changing from V_{DD} to gnd.
  - $V_{in} > V_b$-V_{th} M_1 is off.
  - For lower values of V_{in} assuming M_1 is in saturation:

$$I_D = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_b - V_{in} - V_{TH})^2$$

  - As V_{in} decreases M_1 gets into triode if:

$$V_{DD} - \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_b - V_{in} - V_{TH})^2 R_D = V_b - V_{TH}$$

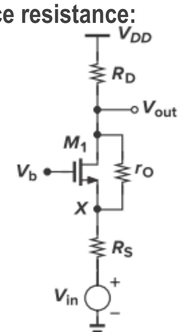
  - If M_1 is in saturation: $V_{out} = V_{DD} - \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_b - V_{in} - V_{TH})^2 R_D$
  - Thus taking derivative and noting that we get: $\frac{\partial V_{TH}}{\partial V_{in}} = \frac{\partial V_{TH}}{\partial V_{SB}} = \eta$

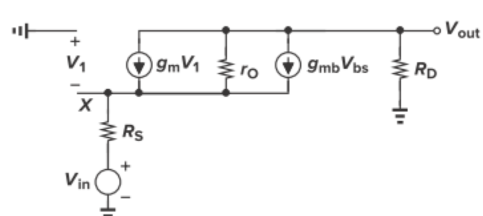
$$\frac{\partial V_{out}}{\partial V_{in}} = -\mu_n C_{ox}\frac{W}{L}(V_b - V_{in} - V_{TH})(-1 - \frac{\partial V_{TH}}{\partial V_{in}})R_D = g_m(1+\eta)R_D$$

Body effect increases the gain.

Common Gate

- Analysis with finite source resistance:





- Current through R_S is V_{out}/R_D. Thus KVL gives: $V_1 - \frac{V_{out}}{R_D}R_S + V_{in} = 0$
- Current through r_o is $-V_{out}/R_D - g_m V_1 - g_{mb} V_1$ thus KVL:

$$r_o\left(\frac{-V_{out}}{R_D} - g_m V_1 - g_{mb} V_1\right) - \frac{V_{out}}{R_D}R_S + V_{in} = V_{out}$$

- Eliminating V_1 we get, $r_o\left[\frac{-V_{out}}{R_D} - (g_m + g_{mb})\left(V_{out}\frac{R_S}{R_D} - V_{in}\right)\right] - \frac{V_{out}}{R_D}R_S + V_{in} = V_{out}$

$$\frac{V_{out}}{V_{in}} = \frac{(g_m + g_{mb})r_o + 1}{r_o + (g_m + g_{mb})r_o R_S + R_S + R_D}R_D$$

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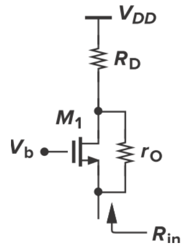
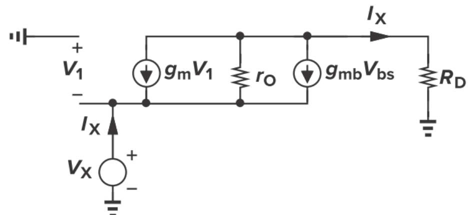
Common Gate

- Input impedance:





- **Current through r_o is $I_X + g_m V_1 + g_{mb} V_1 = I_X - (g_m + g_{mb}) V_X$**
- **Adding up voltages across r_o and R_D we get:**

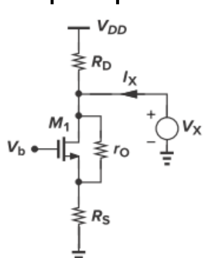
$$R_D I_X + r_o[I_X - (g_m + g_{mb})V_X] = V_X$$

- **Thus, $\frac{V_X}{I_X} = \frac{R_D + r_o}{1 + (g_m + g_{mb}) r_o} \approx \frac{R_D}{(g_m + g_{mb}) r_o} + \frac{1}{g_m + g_{mb}}$ ➔ drain impedance is divide by $(g_m + g_{mb}) r_o$ when seen at the source.**
- **Case-1: R_D is a short then** $\frac{V_X}{I_X} = \frac{r_o}{1 + (g_m + g_{mb}) r_o} = \frac{1}{\frac{1}{r_o} + g_m + g_{mb}}$

Case-2: R_D is open (ideal current source) $\frac{V_X}{I_X} = \infty$

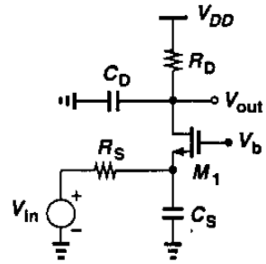
Common Gate

- **Output impedance:** $R_{out} = \{[1 + (g_m + g_{mb}) r_o] R_S + r_o\} || R_D$



- Similar to the case with source degeneration.
- Impedance is greater when source resistance is present.

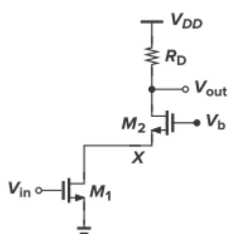
Common Gate-Frequency Response

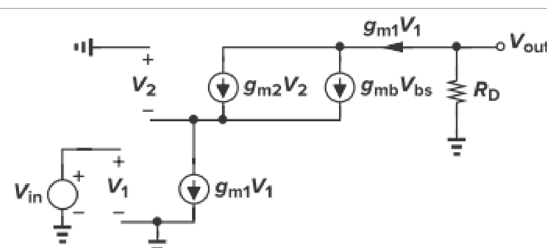


$$\lambda=0 \Rightarrow \frac{V_{out}}{V_{in}}(s) = \frac{(g_m + g_{mb})R_D}{1+(g_m+g_{mb})R_S} \frac{1}{\left(1+\frac{C_S}{g_m+g_{mb}+R_S^{-1}}s\right)(1+R_D C_{DS})}$$

- No Miller multiplication of capacitance as there is no path between input and output other than the transistor.

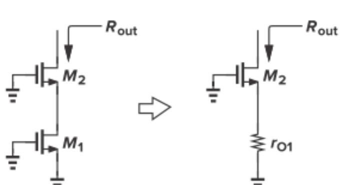
Cascode Stage-Cascaded Triode





Small-signal equivalent circuit with ideal-transistor

- Output Resistance:



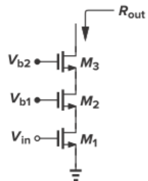
$$R_{out} = [1+(g_{m2}+g_{mb2})r_{o2}]r_{o1} + r_{o2}$$

$$\Longrightarrow R_{out} \approx (g_{m2}+g_{mb2})r_{o2}r_{o1}$$

$$\Longrightarrow R_{out} \approx g_{m2}r_{o2}r_{o1}$$

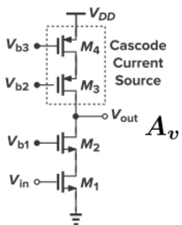
Cascode Stage-Cascaded Triode

- **Triple Cascode ➔**



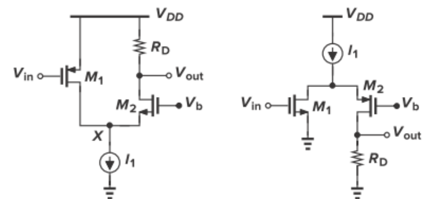
$$\Longrightarrow \; R_{out} \approx g_{m3} g_{m2} r_{o3} r_{o2} r_{o1}$$

- **NMOS Cascode amplifier with PMOS Cascode Load:**



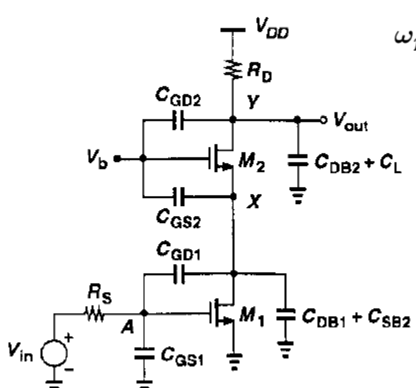
$$A_v \approx g_{m1}[(g_{m2} r_{o2} r_{o1}) || (g_{m3} r_{o3} r_{o4})]$$

- **Folded Cascode:**



Cascode Stage-Frequency Response

High frequency model of a Cascode:



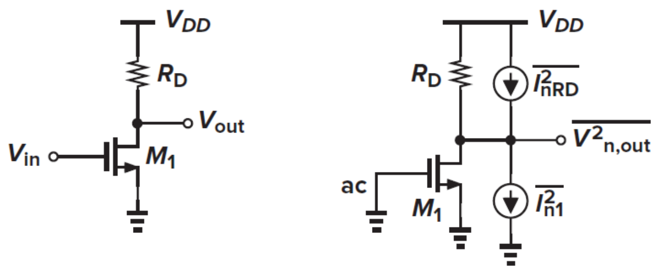
$$\omega_{p,A} = \frac{1}{R_S\left[C_{GS1} + \left(1 + \frac{g_{m1}}{g_{m2} + g_{mb2}}\right) C_{GD1}\right]}$$

Miller effect is not significant when λ=0.

$$\omega_{p,X} = \frac{g_{m2} + g_{mb2}}{2C_{GD1} + C_{DB1} + C_{SB2} + C_{GS2}}$$

$$\omega_{p,Y} = \frac{1}{R_D(C_{DB2} + C_L + C_{GD2})}$$

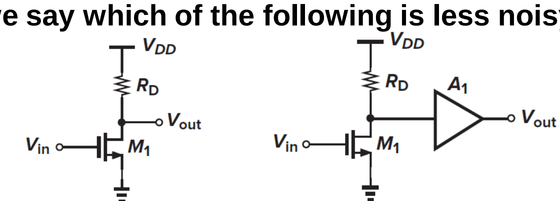
Representation of Noise in Circuits



$$\overline{V_{n,OUT}^2} = \Big(\underbrace{4kT\frac{2}{3}g_m}_{TransistorThermalNoise} + \underbrace{\frac{K}{C_{ox}WL}\frac{1}{f}g_m^2}_{TransistorFlickerNoise} + \underbrace{\frac{4KT}{R}}_{ResistorNoise} \Big) R_D^2$$

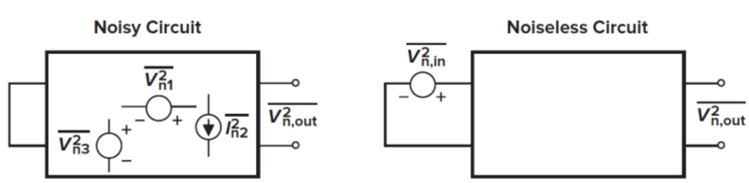
Representation of Noise in circuits

- How do we say which of the following is less noisy?



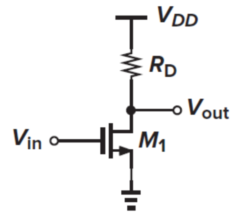
(a) (b)

- If we compare output noise “Circuit (b)” is more noisy
- However “Circuit (b)” amplifies the signal more also ➔ SNR is independent of A_1
- So we



Representation of Noise in circuits

- **Example**



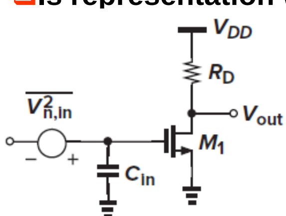
- **Input referred noise is,**

$$\overline{V_{n,in}^2} = \frac{\overline{V_{n,OUT}^2}}{A_v^2}$$

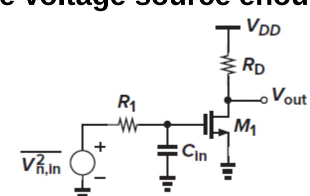
$$\overline{V_{n,in}^2} = 4KT\frac{2}{3g_m} + \frac{K}{C_{ox}WL}\frac{1}{f} + \frac{4KT}{g_m^2 R_D}$$

Representation of Noise in circuits

- **Is representation with just one voltage source enough?**



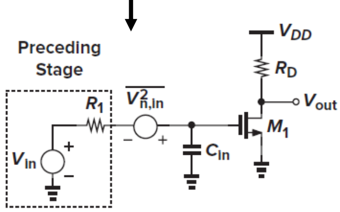
With just noise source



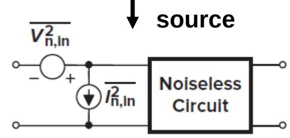
$$\overline{V_{n,out}^2} = \overline{V_{n,in}^2}\left|\frac{1}{1+R_1C_{in}j\omega}\right|^2(g_mR_D)^2$$

$$= \frac{4kT\gamma g_m R_D^2}{R_1^2C_{in}^2\omega^2+1}$$

Output noise diminishes as R_1 increases

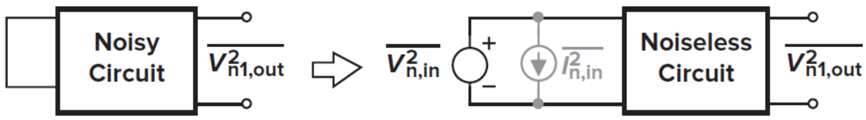


Represent noise both as current and voltage source

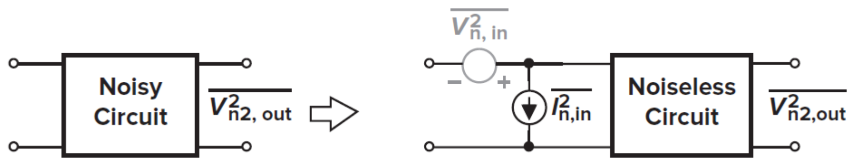


Calculation of Input Referred Noise

- **Zero-Source Impedance (Input Shorted)➔ $\overline{I_{n,in}^2}$ flows through $\overline{V_{n,in}^2}$ and has no effect on the output.**



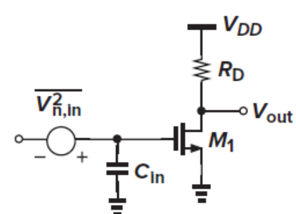
- **Infinite-Source Impedance (Input Opened) ➔ $\overline{V_{n,in}^2}$ has no effect and the output noise is due to only $\overline{I_{n,in}^2}$**



Calculation of Input Referred Noise

- **Example:**
  - **Ignoring Flicker Noise**

$$\overline{V_{n,OUT}^2} = 4kT\frac{2}{3}g_m R_D^2 + 4kTR_D$$



  - **Thus,**

$$\overline{V_{n,in}^2} = 4kT\frac{2}{3g_m} + \frac{4kT}{g_m^2 R_D}$$

  - **If $C_{in} \to \infty$ then the $\overline{V_{n,in}^2}$ gets shorted to ground resulting in zero output voltage. But output noise is present.**
  - **So noise current flowing through C_{in} should generate output voltage**
  - **Thus,**

$$\overline{V_{n,OUT}^2} = \overline{I_{n,in}^2}\left(\frac{1}{C_{in}\omega}\right)^2 g_m^2 R_D^2$$

Calculation of Input Referred Noise

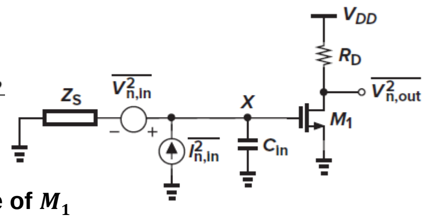
$$4kT\frac{2}{3}g_mR_D^2 + 4kTR_D = \overline{I_{n,in}^2}\left(\frac{1}{C_{in}W}\right)^2 g_m^2R_D^2$$
$$\overline{I_{n,in}^2} = (C_{in}W)^2\frac{4kT}{g_m^2}\left(\frac{2}{3}g_m + \frac{1}{R_D}\right)$$

- **Are we counting noise twice??**
  - **Here, $\overline{V_{n,in}^2}$ and $\overline{I_{n,in}^2}$ represent same noise => they are correlated.**
  - **From previous example,**

$$V_{n,in} = V_{n,M1} + \frac{V_{n,R_D}}{g_m^2R_D^2}$$

and

$$I_{n,in} = C_{in}sV_{n,M1} + \frac{C_{in}sV_{n,R_D}}{g_m^2R_D^2}$$



where,

$V_{n,M1}$ **= gate referred noise voltage of M_1**

$V_{n,RD}$ **= noise voltage of R_D**

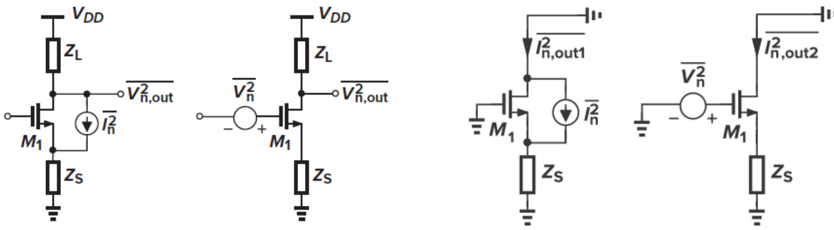
Calculation of Input Referred Noise

- **Since $V_{n,M1}$ and $V_{n,RD}$ appear in both there is a strong correlation between $V_{n,in}$ and $I_{n,in}$**
- **When signals are correlated => use superposition of voltages**
- **When signals are uncorrelated => use superposition of power**
- **Thus,**

$$V_{n,X} = V_{n,in}\frac{\frac{1}{C_{in}s}}{\frac{1}{C_{in}s}+Z_s} + I_{n,in}\frac{\frac{Z_s}{C_{in}s}}{\frac{1}{C_{in}s}+Z_s}$$
$$=> V_{n,X} = \frac{V_{n,in}+I_{n,in}Z_s}{Z_sC_{in}s+1}$$

- **Substituting $V_{n,in}$ and $I_{n,in}$ => $V_{n,X} = V_{n,M1} + \frac{V_{n,R_D}}{g_m^2R_D^2}$**
- **$V_{n,X}$ is independent of Z_s and C_{in}**

LEMMA



Both have equal output impedances ➔ compare the output noise currents.

$$I_{n,out1} = \frac{I_n}{Z_S(g_m + g_{mb} + 1/r_O) + 1} \qquad I_{n,out2} = \frac{g_m V_n}{Z_S(g_m + g_{mb} + 1/r_O) + 1}$$

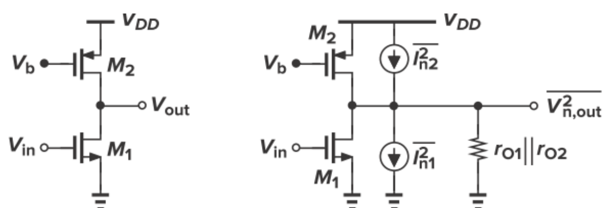
- **The above are equivalent at low frequencies if**

$$V_n^2 = \frac{I_n^2}{g_m^2}$$

  - **Circuit are driven by infinite impedance**

- **Noise source can be transformed from "drain-source" current to a gate series voltage for arbitrary Z_S .**

Voltage Amplification Vs Current Generation

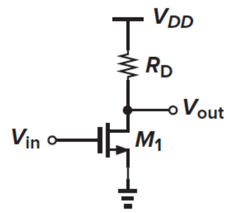


$$\overline{V^2_{n,in}} = 4kT(\gamma g_{m1} + \gamma g_{m2})\frac{1}{g_{m2}^2}$$

$$= 4kT\gamma\left(\frac{1}{g_{m1}} + \frac{g_{m2}}{g_{m1}^2}\right)$$

- **For current sources minimize g_m.**
- **For amplification maximize g_m.**

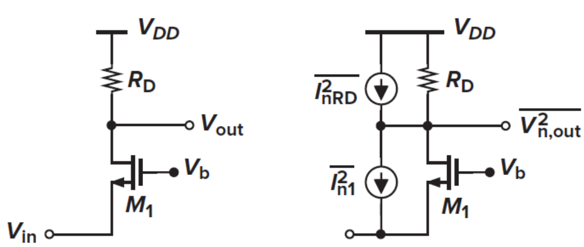
Common Source-Noise Analysis



$$\overline{V_{n,in}^2} = 4kT\left(\frac{2}{3g_m} + \frac{1}{g_m^2 R_D}\right) + \frac{K}{C_{ox}WL}\frac{1}{f}$$

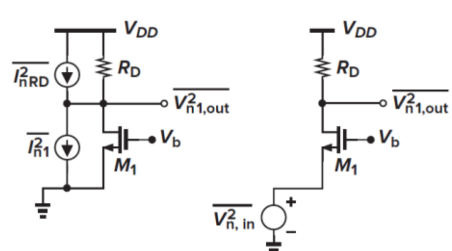
- ❑Minimize noise maximize "g_m"
- ❑Transistor used for voltage amplification
  - ❑=> Minimize noise by maximizing "g_m"
- ❑Transistor used for current generation (or current source)
  - ❑=> Minimize noise by minimizing "g_m"

Common Gate Noise Analysis

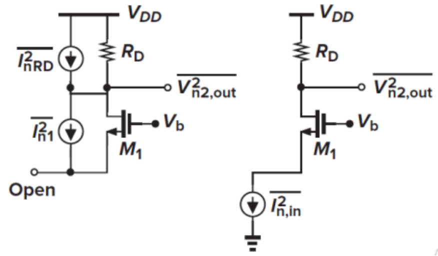


❑**Thus,** $\left(4kT\frac{2}{3}g_m + \frac{4kT}{R_D}\right)R_D^2 = \overline{V_{n,in}^2}(g_m + g_{mb})^2 R_D^2$

$$\overline{V_{n,in}^2} = \frac{4kT\left(2g_m/3 + 1/R_D\right)}{(g_m + g_{mb})^2}$$



Common Gate Noise Analysis



- => $\overline{I_{n1}}^2$ does not contribute? (Why?)
- Thus, $\overline{I_{n,in}}^2 R_D^2 = 4kTR_D$

$$\overline{I_{n,in}}^2 = 4kT/R_D$$

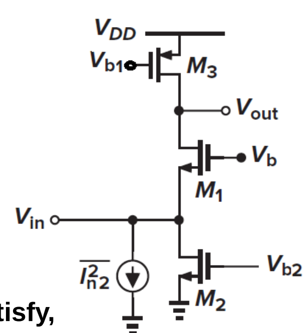
Example

- To compute input referred voltage

  compute $\overline{V_{n,OUT}}^2$ by shorting input to

  ground.
- Thus,

$$\overline{V_{n,OUT}}^2 = 4kT\frac{2}{3}(g_{m1} + g_{m3})(r_{01}||r_{03})^2$$



- Thus, input referred noise voltage must satisfy,

$$\overline{V_{n,in}}^2 (g_{m1} + g_{mb1})^2 (r_{01}||r_{03})^2 = 4kT\frac{2}{3}(g_{m1} + g_{m3})(r_{01}||r_{03})^2$$

$$=> \overline{V_{n,in}}^2 = 4kT\frac{2}{3}\frac{g_{m1} + g_{m3}}{(g_{m1} + g_{mb1})^2}$$

Example (contd.)

- To calculate input referred current we open the input and compute corresponding output noise.
- Thus,

$$\overline{V_{n,OUT}}^2 = (\overline{I_{n2}}^2 + \overline{I_{n3}}^2)R_{OUT}^2 = \overline{I_{n,in}}^2 R_{OUT}^2$$

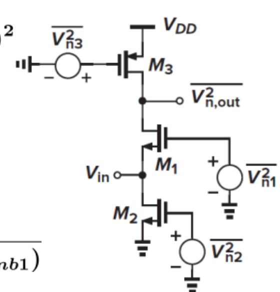
$$=> \overline{I_{n,in}}^2 = (\overline{I_{n2}}^2 + \overline{I_{n3}}^2)$$

$$=> \overline{I_{n,in}}^2 = 4kT\frac{2}{3}(g_{m2} + g_{m3})$$

Noise in Single Stage Amplifiers

- "$1/f$" noise for common gate topology :-
  - Short input to ground:

$$\overline{V_{n,OUT}}^2 = \frac{1}{C_{ox}f}\left[\frac{g_{m1}^2 K_N}{(WL)_1} + \frac{g_{m3}^2 K_P}{(WL)_3}\right](r_{01}||r_{03})^2$$



  - Gain from input to output is,

$$A_v = (g_{m1} + g_{mb1})(r_{01}||r_{03})^2$$

$$\overline{V_{n,in}}^2 = \frac{1}{C_{ox}f}\left[\frac{g_{m1}^2 K_N}{(WL)_1} + \frac{g_{m3}^2 K_P}{(WL)_3}\right]\frac{1}{(g_{m1} + g_{mb1})}$$

  - With input open output noise voltage is given by,

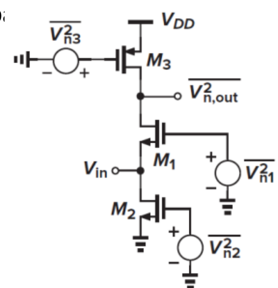
$$\overline{V_{n,OUT}}^2 = \frac{1}{C_{ox}f}\left[\frac{g_{m2}^2 K_N}{(WL)_2} + \frac{g_{m3}^2 K_P}{(WL)_3}\right]R_{OUT}^2$$

Noise in Single Stage Amplifiers

- Why is $\overline{V_{n1}}^2$ noise not considered?
  - Anything to do with degeneration?
  - Thus,

$$\overline{I_{n,in}^2} = \frac{1}{C_{ox}f}\left[\frac{g_{m2}^2 K_N}{(WL)_2} + \frac{g_{m3}^2 K_P}{(WL)_3}\right]$$

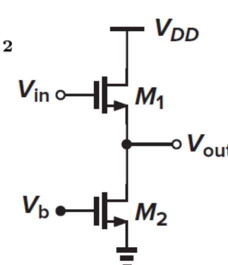
  - If output impedance of M_2 is small and comp
    we get the same expression
  - Try to analyze it with the following circuit



Source Follower

- Now,

$$\overline{V_{n,OUT}^2} = (\overline{I_{n,M1}^2} + \overline{I_{n,M2}^2})\left[\frac{1}{g_{m1}}\bigg|\bigg|\frac{1}{g_{mb1}}\bigg|\bigg|r_{01}\bigg|\bigg|r_{02}\right]^2$$



- Input impedance of common source and source follower is very high for a large bandwidth
- Input referred noise voltage is enough to represent the input referred noise
- Voltage gain of Common Drain or Source Follower is $A_v = \dfrac{\frac{1}{g_{mb1}}\big|\big|r_{01}\big|\big|r_{02}}{\frac{1}{g_{mb1}}\big|\big|r_{01}\big|\big|r_{02} + \frac{1}{g_{m1}}}$

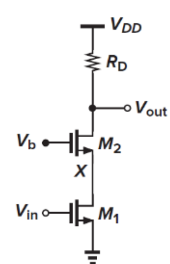
$$\overline{V_{n,in}^2} = \frac{\overline{V_{n,OUT}^2}}{A_v^2} \qquad \overline{V_{n,in}^2} = 4kT\frac{2}{3}\left(\frac{1}{g_{m1}} + \frac{g_{m2}}{g_{m1}^2}\right)$$

- Thus,

Source Follower – Noise Analysis

- Isn't this similar to the noise voltage of common source. Do you know why?
  - How do you compute input referred noise?
    - By Removing inputs i.e. ac ground for inputs
    - Don't CS and Source follower look similar when input ac grounded.
  - If both CS and Common Drain have same input referred noise which circuit is more desirable
    - CS provides signal gain
    - Source follower's gain < 1

Cascode Amplifier – Noise Analysis

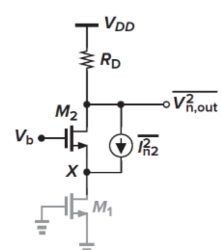


- At low frequency noise currents of M_1 and M_2 flow through RD and noise is thus similar to CS stage,

$$\overline{V_{n,in,M1\&R_D}^2} = 4kT\left(\frac{2}{3g_{m1}} + \frac{1}{g_{m1}^2 R_D}\right)$$

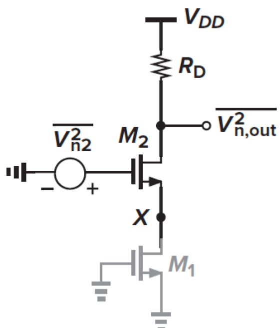
- Where $1/f$ noise is ignored

- What about noise of M_2?
  - If $\lambda = 0$ for M_1 can noise current flow to the output?
    - No.? Why?
  - If $\lambda \neq 0$ for M_1 but output impedance of M_1 is high what happens to the noise current and consequent output noise voltage



Cascode Amplifier – Noise Analysis

- Can you use the lemma to transform above circuit to the following:



Can you use principle of degeneration to show that $\overline{V_{n,OUT,M2}^2}$ is really small?